



Title of the Invention

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Semiconductor Integrated Circuit

Background of the Invention

The present invention relates to a voltage converter which lowers an external supply voltage within a semiconductor integrated circuit chip.

Reduction in the geometries of devices such as bipolar or MOS transistors has been <sup>accompanied by a</sup> ~~attended with~~ lowering in the breakdown voltages of the devices, which has made it <sup>necessary</sup> ~~inevitable~~ to lower the operating voltage of small geometry devices with an integrated circuit <sup>correspondingly</sup>. From the viewpoint of users, however, a single voltage source of <sup>for example which is</sup> 5 V <sup>easy</sup> ~~to~~ use is desirable. As an expedient for meeting such different requests of IC manufacturers and the users, it is considered <sup>to be necessary</sup> ~~to~~ lower the external supply voltage  $V_{CC}$  within a chip and to operate the small geometry devices with the lowered voltage  $V_L$ .

Figure 1 shows an example of <sup>such an</sup> ~~the~~ expedient, in which the circuit A' of the whole chip 10 including, e. g., an input/output interface circuit is operated with the internal supply voltage  $V_L$  lowered by a voltage converter 13.

Figure 2 shows an integrated circuit disclosed in ~~Japanese Patent Application No. 56-57143 (Japanese Laid-open Patent Application No. 57-172761)~~. The small

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geometry devices are employed for a circuit A determining the substantial density of integration of the chip 10, and are operated with the voltage  $V_L$  obtained by lowering the external supply voltage  $V_{CC}$  by means of a voltage converter 13. On the other hand, devices of comparatively large geometries are employed for a driver circuit B including, e. g., an input/output interface which does not greatly contribute to the density of integration, ~~which~~ <sup>which</sup> and they are operated by applying  $V_{CC}$  thereto. Thus, a large-scale integrated circuit (hereinbelow, termed "LSI") which operates with  $V_{CC}$  when viewed from outside the chip becomes possible.

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However, when such <sup>an</sup> integrated circuit is furnished with the voltage converter, an inconvenience is involved in an aging test which is performed after the final fabrication step of the integrated circuit.

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The <sup>terminology</sup> "aging test" ~~is such that~~, after the final fabrication step of the integrated circuit, <sup>63</sup> ~~voltages~~ <sup>during which</sup> higher than in an ordinary operation are intentionally applied to the respective transistors in the circuit, ~~thereby~~ to test the integrated circuit ~~which is liable~~ <sup>for</sup> break down due to an inferior gate oxide film.

The aforementioned voltage converter in Japanese Patent Application No. 56-57143 functions to feed the predetermined voltage. Therefore, the circuit fed with

the supply voltage by the voltage converter cannot be subjected to the aging test.

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In order to solve this problem, an invention disclosed in <sup>U.S. Patent 4,482,985</sup> ~~Japanese Patent Application No. 56-168698~~ has previously been made, but it has had difficulty in the performance for actual integrated circuits. As illustrated in Figures 2 to 6 in the patent application, according to <sup>that</sup> ~~the~~ cited invention, an internal voltage increases up to an aging point rectilinearly or with one step of change as an external supply voltage increases. Accordingly, the internal voltage changes greatly with the change of the external supply voltage. This has led to the disadvantage that the breakdown voltage margins of small geometry devices in an ordinary operation become small.

#### Summary of the Invention

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An object of the present invention is to further advance the invention of <sup>disclosed in U.S. Patent 4,482,985</sup> ~~Japanese Patent Application No. 56-168698~~ referred to above, and to provide a voltage converter <sup>(C)</sup> ~~which~~ can widen the margins of the breakdown voltages of small geometry devices in an ordinary operation and which affords sufficient voltages in an aging test.

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The present invention consists in that the output voltage of <sup>the</sup> ~~a~~ voltage converter is set at a voltage suitable for the operations of small geometry devices against the change of an external supply voltage when a semiconductor

integrated circuit is in its ordinary operation region, and at an aging voltage when the ordinary operation region is exceeded.

To this end, according to the voltage converter of the present invention, when the external supply voltage has been changed from the lower limit value of the ordinary operation range thereof to the aging operation point thereof, the output voltage of the voltage converter changes up to the aging voltage without exhibiting a constant changing rate versus the change of the external supply voltage.

#### Brief Description of the Drawings

Figures 1 and 2 show semiconductor integrated circuits each having a voltage converter.

Figures 3 and 5 show basic circuits each of which constitutes a device embodying the present invention.

Figures 4 and 6 show the characteristics of the circuits in Figures 3 and 5, respectively.

Figures 7, 9 and 11 show devices embodying the present invention.

Figures 8, 10 and 12 show the characteristics of the circuits in Figures 7, 9 and 11, respectively.

Figures <sup>13 (A) to (C), 14 (A) to (C) formation of the</sup> ~~13~~ and ~~14~~ <sup>^</sup> <sub>^</sub> <sup>^</sup> <sub>^</sub> show the circuit of Figure 3 in practicable forms.

Figure 15 shows the characteristic in Figure 4

more specifically.

Figure 16 shows another practicable example of the circuit in Figure 3.

Figure 17 shows the characteristic in Figure 8 concretely.

Figure 18 shows a circuit for producing the characteristic in Figure 17.

Figure 19 shows the characteristic in Figure 8 concretely.

Figure 20 shows a circuit for producing the characteristic in Figure 19.

Figure 21 shows the characteristic in Figure 10 concretely.

Figure 22 shows a circuit for producing the characteristic in Figure 21.

Figure 23 shows a characteristic in another embodiment of the present invention.

Figure 24 shows a circuit for producing the characteristic in Figure 23.

Figure 25 shows the characteristic in Figure 12 concretely.

Figure 26 shows a circuit for producing the characteristic in Figure 25.

Figure 27 shows a practicable example of the circuit in Figure 26.

Figure 28 shows the actual characteristics of the circuit in Figure 27.

Figure 29(A) shows a gate signal generator for use in an embodiment of the present invention.

Figure 29(B) shows a time chart of the circuit in Figure 29(A).

Figure 30 shows a protection circuit which connects the circuit of Figure 29(A) with the circuit of Figure 16, 18, 20, 22, 24 or 26.

Figure 31 shows a practicable circuit of an inverter for use in the circuit of Figure 29(A).

Figure 32 shows a practicable circuit of an oscillator for use in the circuit of Figure 29(A).

Figure 33 shows an example of a buffer circuit for the output of the circuit shown in Figure 16, 18, 20, 22, 24 or 26.

Figure 34 shows the characteristics of the circuit in Figure 33.

Figures 35, 36 and 37 show other examples of buffer circuits, respectively.

Figure 38 shows a time chart of the circuit in Figure 37.

Figure 39 shows a practicable example of the circuit in Figure 3.

Figure 40 shows an example of a buffer circuit.

Figure 41 shows the characteristics of the circuit in Figure 40.

Description of the Preferred Embodiments

Voltage converter circuit forms for affording various output characteristics versus an external supply voltage  $V_{CC}$ , as well as practicable examples thereof, will be first described, followed by practicable embodiments on a method of feeding power to the voltage converter and on a buffer circuit for the voltage converter well suited to drive a high load.  $\Delta$

Figures 3 and 5 show basic circuits each of which constitutes the voltage converter of the present invention.

In the circuit of Figure 3, a resistance  $R_3$  in Figure <sup>73</sup> ~~34~~ of <sup>U.S. Patent 4,482,985</sup> ~~Japanese Patent Application No. 56-168698~~ already filed is made variable, and a transistor Q is employed in order to enhance a current driving ability for a load to which an output voltage  $V_L$  is applied. Here, the control terminal voltage  $V_G$  of the transistor Q has a characteristic which changes versus the change of an external supply voltage  $V_{CC}$  and which is the output voltage of a reference voltage generator REF. More specifically, as illustrated in Figure 4, in a case where the external supply voltage  $V_{CC}$  is gradually increased from 0 (zero) V, the voltage  $V_G$  rises abruptly when a certain voltage  $V_P$  has been reached, so that the transistor Q

turns "on". For  $V_{CC}$  not smaller than  $V_P$ , Q continues to turn "on". Therefore, the effective impedance of the whole basic circuit BL <sup>decreases</sup> ~~lowers~~, and the ratio thereof with the effective impedance R changes, so that the voltage  $V_L$  becomes a straight line of different slope for  $V_{CC}$  not smaller than  $V_P$  as shown in Figure 4. Here in Figure 4, the example is illustrated in which  $V_G$  rises abruptly from 0 V to a certain voltage for  $V_{CC}$  not smaller than  $V_P$ . However, it is also allowed to adopt a characteristic in which, in case of changing  $V_{CC}$  from 0 V,  $V_G$  rises gradually from 0 V and becomes, at the point  $V_P$ , a voltage level to turn "on" the transistor Q. Regarding the example in which  $V_G$  rises abruptly at and above the certain voltage  $V_{CC}$ , the reference voltage generator can be realized by the cascade connection of devices having rectification characteristics as taught in <sup>U.S. Patent 4,482,985</sup> ~~Japanese Patent Application No. 56-168698~~. Regarding the example in which  $V_G$  rises gradually, the reference voltage generator can be realized by a simple resistance divider circuit. In Figure 4, the coefficient of  $V_L$  relative to  $V_{CC}$  can be changed at will by the designs of the resistance and the transistor Q.

Figure 5 shows another example which employs the same basic circuit BL as in Figure 3. Whereas the example of Figure 3 derives  $V_L$  from the  $V_{CC}$  side, this example



derives  $V_L$  from the ground side. When the characteristic of the output voltage  $V_G$  from the reference voltage generator is set in advance so that the transistor Q may turn "on" at  $V_{CC}$  not smaller than  $V_P$ ,  $V_L$  is determined by the effective impedance of the whole basic circuit BL and the effective impedance R, and hence,  $V_L$  becomes as shown in Figure 6.

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b While Figures 3 and 5 have exemplified the transistors as being ~~the~~ MOS transistors, bipolar transistors may <sup>if desired</sup> well be used. Particularly in a case where ~~the~~ whole chips are constructed of MOS transistors in the examples of Figures 1 and 2, it is usually easier to design them when the circuits of Figures 3 and 5 are constructed of MOS transistors. In a case where the whole chips are of bipolar transistors, it is more favorable to use bipolar transistors. It is sometimes the case, however, that the chip includes both MOS transistors and bipolar transistors. It is to be understood that, in this case, the MOS transistor or/and the bipolar transistor can be used for the circuit of Figure 3 or Figure 5 in accordance with an intended application. In addition, although the examples of Figures 4 and 6 have been mentioned as the characteristics of the circuit REF, these examples are not especially restrictive, but the characteristic of the circuit REF may be set according to the purpose

of the design of  $V_L$ .

Now, the voltage converter based on the circuit of Figure 3 will be described. Figures 7 and 8 illustrate an example in which the basic circuits BL numbering  $k$  are connected in parallel with the effective impedance  $R$  of the circuit of Figure 3. The circuits REF in the respective basic circuits BL are set so that  $BL_0$  may first turn "on" at  $V_{P0}$ ,  $BL_1$  may subsequently turn "on" at  $V_{P1}$ , and  $BL_k$  may lastly turn "on" at  $V_{Pk}$ . The transistors in the respective circuits BL are designed so that the coefficients of the changes of the respective voltages  $V_L$  versus the voltage  $V_{CC}$  may be varied. As  $V_{CC}$  increases more, impedances are successively added in parallel with the impedance  $R$ , so that the entire characteristic of  $V_L$  becomes concave for  $V_{CC}$  not smaller than  $V_{P0}$ .

The coefficients of the changes are varied for the following reason. For example, in a case where the aging operation points are  $V_{P2}$ ,  $V_{P3}$ , ..... and  $V_{Pk}$  and where the aging voltages of circuits to be fed with the supply voltages by the voltage converter are  $V_{L2}$ ,  $V_{L3}$ , ..... and  $V_{Lk}$ , the transition is smoothed when the first aging operation point shifts to the next one.

The present circuit is a circuit which is practical in points of the operating stability of the ordinary operation and an effective aging for the system of Figure 2.

By way of example, the  $V_{CC}$  operation point in the ordinary operation is set at a point at which  $V_L$  changes versus  $V_{CC}$  as slightly as possible, that is, the coefficient of change is the smallest, in order to achieve a stable operation, whereas the  $V_{CC}$  operation point in the aging test is set at a point at which the coefficient of change is great, in order to approximately equalize the stress voltage conditions of a transistor of large geometry and a transistor of small geometry as described in Japanese Patent Application No. 56-168698. More concretely, in case of using only  $BL_0$  and  $BL_1$  in the circuit of Figure 7, the coefficient of change in Figure 8 may be made small between the lower limit voltage  $V_{P0}$  (e. g., 2 - 3 V) and the upper limit voltage  $V_{P1}$  (e. g., 6 V), to set the ordinary operation point (e. g., 5 V) concerning  $V_{CC}$  for the ordinary operation range in this section, while the coefficient of change may be made great between  $V_{P1}$  and  $V_{P2}$  (e. g., 7 - 9 V), to set the aging operation point (e. g.,  $V_{CC} = 8$  V) in this section. The ordinary operation range is solely determined by ratings, and it is usually set at  $5 \pm 0.5$  V. It is to be understood that, for some purposes of designs, the operation voltage points and the aging voltage points can be set at any desired  $V_{CC}$  points by employing the basic circuits  $BL_2$ ,  $BL_3$  ... more. When the more circuits BL are used, the

$V_L$  characteristic can also be made smoother versus  $V_{CC}$ , so that the operation of the internal circuit can be stabilized more. Further, since the  $V_{CC}$  voltage is high in the aging test, it is effective to construct the voltage converter itself by the use of high breakdown voltage transistors. To this end, the voltage converter may be constructed of transistors of large geometry in the system of Figure 2 by way of example.

Figures 9 and 10 show an example in which the basic circuits BL are connected in parallel on the ground side. As stated before, when the respective circuits BL are properly designed, the characteristic of the whole  $V_L$  can be made convex relative to  $V_{CC}$ . This characteristic is effective for protecting the circuit A' from any overvoltage  $V_L$  in the system of Figure 1 by way of example. This achieves the advantage that, in case of measuring the  $V_{CC}$  voltage margin of the whole chip, a sufficiently high voltage  $V_{CC}$  can be applied without destroying small geometry devices.

In some uses, it is also possible that the circuits of Figures 7 and 9 coexist. By way of example, the ordinary operation point is set at a point at which the coefficient of change is small, and the aging operation point is set at a point at which the coefficient of change is great. These are realized by  $BL_0$  and  $BL_1$  in

the circuit of Figure 7. Further, in order to make the coefficient of change small again at and above the  $V_{CC}$  point of the aging condition to the end of preventing the permanent breakdown of devices, the basic circuits BL other than  $BL_0$  are connected so as to operate in parallel with the latter as in the circuit form of Figure 9. This makes it possible to design a circuit in which the devices are difficult to break down at and above the  $V_{CC}$  point of the aging condition.

Thus, even when the supply voltage has been made abnormally high erroneously by way of example, the breakdown of the devices can be prevented.

Figures 11 and 12 show an example in which a basic circuit BL' is connected in parallel with the circuit of Figure 3, whereby the changing rate of  $V_L$  is made negative at and above  $V_P'$  which is a certain value of  $V_{CC}$ . More specifically, when  $V_{CC}$  is increased, the transistor Q first turns "on" while the output voltage  $V_G$  of the reference voltage generator 1 in the basic circuit BL is not lower than  $V_P$ , so that the gradient of  $V_L$  versus  $V_{CC}$  decreases. A reference voltage generator 2 is designed so that a transistor Q' in the basic circuit BL' may subsequently turn "on" at the certain  $V_{CC}$  value, namely,  $V_P'$ . In addition, the conductance of Q' is designed to be sufficiently higher than that of Q. Then,

the  $V_L$  characteristic after the conduction of the transistor  $Q'$  is governed by the characteristic of  $BL'$ , so that  $V_L$  comes to have the negative gradient as shown in Figure 12.

The merit of the present circuit is that, when the aforementioned point at which  $V_L$  lowers is set at or below the breakdown voltages of small geometry devices, these small geometry devices are perfectly protected from breakdown even when the voltage  $V_{CC}$  has been sufficiently raised. For example, a measure in which the output voltage  $V_L$  lowers when a voltage higher than the external supply voltage  $V_{CC}$  at the aging point has been applied is especially effective because any voltage exceeding the aging point is not applied to the devices.

It is to be understood that an external instantaneous voltage fluctuation can also be coped with.

Obviously, the circuit of Figure 5 can afford any desired  $V_L$  characteristic by connecting the basic circuit  $BL'$  in parallel as in the example of Figure 3.

While, in the above, the conceptual examples of the voltage converters have been described, practicable circuit examples based on these concepts will be stated below.

Figure 13 shows <sup>a</sup> practicable example of the circuit of Figure 3 which employs a bipolar transistor. A voltage regulator circuit CVR is, for example, a cascade

connection of Zener diodes or ordinary diodes the terminal voltage of which becomes substantially constant. (A) indicates a well-known voltage regulator. This is described in detail in "Denpa-Kagaku (Science of Electric Wave)", February 1982, p. 111 or "Transistor Circuit Analysis", by Joyce and Clarke, Addison-Wesley Publishing Company, Inc., p. 207. Since, however,  $V_L$  is a fixed voltage in this condition, a resistance  $r$  is connected in series with CVR as shown in (B). Thus,  $V_L$  comes to have a slope relative to  $V_{CC}$  as shown in (C).

Figure 14 shows another embodiment. (A) indicates a well-known voltage regulator which employs an emitter follower. Since  $V_L$  is also a fixed voltage, a resistance  $r$  is used in (B) as a measure of solution. Thus, a characteristic as shown in (C) is afforded.

These examples of Figures 13 and 14 are especially suited to the system as shown in Figure 1. In Figure 1, usually a great current flows through the circuit associated with the input/output interface. Therefore, a high current driving ability is required of the voltage converter correspondingly. Obviously, the voltage converter constructed of the bipolar transistor is suited to this end.

Next, there will be explained practicable examples in which voltage converters are constructed of MOS transistors on the basis of the circuits of Figures 3, 7, 9 and 11.

Figure 15 shows a concrete example of the characteristic of Figure 4 in which  $V_L$  is endowed with a slope  $m$  for  $V_{CC}$  of and above a certain specified voltage  $V_0$ . Since the change of  $V_L$  decreases for the voltage not smaller than  $V_0$ , the breakdown of small geometry devices becomes difficult to occur to that extent.

$V_L = V_{CC}$  is held for  $V_{CC}$  smaller than  $V_0$ , for the following reason. In general, MOSTs have the operating speeds degraded by lowering in the threshold voltages thereof as the operating voltages lower. To the end of preventing this drawback, it is desirable to set the highest possible voltage on a lower voltage side such as  $V_{CC}$  smaller than  $V_0$ . That is,  $V_L$  should desirably be equal to  $V_{CC}$ .

Figure 16 shows an embodiment of a practicable circuit DCV therefor, which corresponds to a practicable example of the circuit of Figure 3.

The features of the present circuit are that the output voltage  $V_L$  is determined by the ratio of the conductances of MOS transistors  $Q_0$  and  $Q_1$ , and that the conductance of the MOS transistor  $Q_1$  is controlled by  $V_L$ .  
*the output voltage*  
 c by  $V_L$

With the present circuit, letting the gate voltage  $V_G$  of  $Q_0$  be  $V_{CC} + V_{th(0)}$  (where  $V_{th(0)}$  denotes the threshold voltage of the MOST  $Q_0$ ), the control starting voltage  $V_0$



and the slope  $\underline{m}$  are expressed as follows:

$$V_O = \sum_{i=1}^n V_{th(i)} + V_{th(l)}$$

$$m = \left\{ 1 + \sqrt{\beta(l)/\beta(0)} \right\}^{-1}$$

Here,  $\beta(0)$  and  $\beta(l)$  denote the channel conductances of  $Q_0$  and  $Q_l$ ,  $V_{th(i)}$  ( $i = 1 - n$ ) and  $V_{th(l)}$  denote the threshold voltages of the MOS transistors  $Q_i$  ( $i = 1 - n$ ) and  $Q_l$ , and  $\underline{n}$  denotes the number of stages of  $Q_i$ .

Accordingly,  $V_O$  and  $\underline{m}$  can be varied at will by  $\underline{n}$ ,  $V_{th(i)}$ ,  $V_{th(l)}$  and  $\beta(l)/\beta(0)$ . It has been stated before that  $V_L = V_{CC}$  is desirable for  $V_{CC}$  smaller than  $V_O$ . In this regard, for  $V_{CC}$  smaller than  $V_O$ ,  $V_L$  is determined by  $V_O$  because  $Q_l$  is "off". Therefore, the voltage  $V_G$  of  $Q_0$  must be a high voltage of at least  $V_{CC} + V_{th(0)}$ .

In order to simplify the computation and to facilitate the description, the circuit of Figure 16 is somewhat varied from an actual circuit. As a practical circuit, as shown in Figure 27 to be referred to later, a transistor of similar connection ( $Q_{S(1.6)}$  in Figure 27) needs to be further connected between the  $n$ -th one of the transistors connected in cascade and the ground. That is, a kind of diode connection is made toward the ground. With this measure, when  $V_{CC}$  has been varied from the high

voltage side to the low voltage side, the nodes of the transistors connected in cascade are prevented from floating states to leave charges behind. For the sake of the convenience of the description, the transistor of this measure shall be omitted in the ensuing embodiments.

Figure 17 shows a characteristic in which, when the external supply voltage  $V_{CC}$  changes between the lower limit value  $V_O$  and upper limit value  $V_O'$  of the ordinary operation range, the slope  $\underline{m}$  of the output voltage  $V_L$  is small, and a slope  $m'$  which corresponds to the external supply voltage greater than  $V_O'$  is made steeper than  $\underline{m}$ .

Figure 18 shows an example of a circuit for producing the characteristic of Figure 17.

These correspond to a practicable form of the example of Figures 7 and 8.

The feature of the present circuit is that, between the terminals 1 and 2 of the circuit DCV shown in Figure 16, a circuit DCV2 similar to DCV1 is added, whereby the conductance of a load for DCV1 is increased at and above  $V_O'$  so as to increase the slope of  $V_L$ .

With the present circuit, the second control starting voltage  $V_O'$  is expressed by:

$$V_O' = V_O + \left\{ \sum_{i=1}^n V_{th(i)} + V'_{th(l)} \right\} / (1 - m)$$

In addition, the slope  $m'$  is determined by the ratio between the sum of the conductances of the MOS transistors  $Q_0$  and  $Q'_l$  and the conductance of the MOS transistor  $Q_l$ . Here,  $V'_{th(i)}$  ( $i = 1 - n'$ ) and  $V'_{th(l)}$  denote the threshold voltages of the MOS transistors  $Q'_i$  ( $i = 1 - n'$ ) and  $Q'_l$ , respectively.

Accordingly,  $V'_0$  and  $m'$  can be varied at will by  $n$ ,  $n'$ ,  $\beta(l)$ ,  $\beta'(l)$ ,  $V_{th(i)}$ ,  $V_{th(l)}$ ,  $V'_{th(i)}$  and  $V'_{th(l)}$ . Here,  $\beta'(l)$  denotes the channel conductance of the MOS transistor  $Q'_l$ .

This circuit has the ordinary operation range between the lower limit value  $V_0$  and the upper limit value  $V'_0$ , and is effective when the aging point has a value larger than  $V'_0$ . That is, since the slope  $m$  is small in the ordinary operation region, margins for the breakdown voltages of small geometry devices are wide, and power consumption does not increase. Here, the slope  $m'$  for the external supply voltage higher than the ordinary operation region is set for establishing a characteristic which passes an aging voltage (set value).

In an example illustrated in Figure 19, a characteristic in which the slope of  $V_L$  becomes  $m''$  greater than  $m'$  when the external supply voltage  $V_{CC}$  has reached  $V_0''$  is further added to the characteristic shown in Figure 17.

Figure 20 shows an example of a practicable circuit

therefor. These correspond to a concrete form of the example of Figures 7 and 8. The feature of the present circuit is that circuits DCV2 and DCV3 similar to the circuit DCV1 are added between the terminals 1 and 2 of the circuit shown in Figure 16, whereby the conductance of the load for DCV1 is successively increased so as to increase the slope of  $V_L$  in two stages at the two points  $V_O'$  and  $V_O''$ .

With the present circuit, the second and third control starting voltages  $V_O'$  and  $V_O''$  are respectively expressed by:

$$V_O' = V_O + \left\{ \sum_{i=1}^{n'} V'_{th(i)} + V'_{th(l)} \right\} / (1 - m)$$

$$V_O'' = V_O' + \left\{ \sum_{i=1}^{n''} V''_{th(i)} + V''_{th(l)} - \sum_{i=1}^{n'} V'_{th(i)} - V'_{th(l)} \right\} / (1 - m')$$

Here,  $V''_{th(i)}$  ( $i = 1 - n''$ ) and  $V''_{th(l)}$  denote the threshold voltages of the MOS transistors  $Q''_i$  ( $i = 1 - n''$ ) and  $Q''_l$ , respectively. Besides, the slope  $m'$  is determined by the ratio between the sum of the conductances of the MOS transistors  $Q_O$  and  $Q'_l$  and the conductance of the MOS transistor  $Q_l$ , and the slope  $m''$  by the ratio between the sum of the conductances of the MOS transistors  $Q_O$ ,  $Q'_l$  and  $Q''_l$  and the conductance of the MOS transistor  $Q_l$ .

Accordingly,  $V_O'$  and  $m'$  can be varied at will by  $n$ ,  $n'$ ,  $\beta(0)$ ,  $\beta(l)$ ,  $\beta'(l)$ ,  $V_{th(i)}$ ,  $V_{th(l)}$ ,  $V'_{th(i)}$  and  $V'_{th(l)}$ ,

while  $V''_0$  and  $m''$  by  $n, n', n'', \beta(0), \beta(l), \beta'(l), \beta''(l), V_{th(i)}, V_{th(l)}, V'_{th(i)}, V'_{th(l)}, V''_{th(i)}$  and  $V''_{th(l)}$ . Here,  $\beta''(l)$  denotes the channel conductance of  $Q''_l$ .

This circuit is effective when the ordinary operation range extends ~~from~~ the lower limit value  $V_0$  and the upper limit value  $V_0'$ , and aging tests are carried out in the two sections of the external supply voltage  $V_{CC} \geq V_0''$  and  $V_0' < V_{CC} < V_0''$ . The aging tests in the two sections consist of the two operations; aging for a short time, and aging for a long time. The former serves to detect a defect occurring, for example, when an instantaneous high stress has been externally applied, while the latter serves to detect a defect ascribable to a long-time stress.

Figure 21 shows an example wherein, when the external supply voltage  $V_{CC}$  is greater than  $V_0'$ , the slope  $m'$  of the voltage  $V_L$  is set at  $m > m'$  under which the output voltage  $V_L$  follows up the external supply voltage  $V_{CC}$ .

Figure 22 shows an embodiment of a practicable circuit therefor. These correspond to a concrete form of the example of Figures 9 and 10. The feature of the present circuit is that a circuit DCV2 similar to DCV1 is added between the terminal 2 and ground of the circuit shown in Figure 16, whereby the conductance of a load for the transistor  $Q_0$  is increased at  $V_0'$  so as to decrease the slope of  $V_L$ .

With the present circuit, the second control starting voltage  $V_O'$  is expressed by:

$$V_O' = -\frac{1-m}{m} V_O + \left\{ \sum_{i=1}^{n'} V'_{th(i)} + V'_{th(l)} \right\} / m$$

In addition, the slope  $m'$  is expressed by the ratio between the conductance of  $Q_O$  and the sum of the conductances of  $Q_L$  and  $Q'_L$ .

Accordingly,  $V_O'$  and  $m'$  can be varied at will by  $n$ ,  $n'$ ,  $\beta(0)$ ,  $\beta(l)$ ,  $\beta'(l)$ ,  $V_{th(i)}$ ,  $V_{th(l)}$ ,  $V'_{th(i)}$  and  $V'_{th(l)}$ .

This circuit is applicable to devices of lower breakdown voltages. Usually, when the breakdown voltages of devices are low, the output voltage  $V_L$  of the ordinary operation region ( $V_O < V_{CC} < V_O'$ ) may be suppressed to a low magnitude. In some cases, however, the magnitude of  $V_L$  cannot be lowered because the operating speeds of a circuit employing small geometry devices and a circuit employing large geometry devices are matched. In such cases, the slope  $m_a$  of the output voltage  $V_L$  in the ordinary operation region is made greater than  $\underline{m}$  indicated in Figure 17 so as to bring  $V_L$  closer to the change of the external supply voltage. When the ordinary operation region has been exceeded, the slope of  $V_L$  is decreased in order for the aging operation point to be passed. Thus, the magnitude of the output voltage  $V_L$

can be raised near to the withstand voltage limit of the devices within the range of the ordinary operation region, and the operating speed of the circuit employing the small geometry devices can be matched with that of the circuit employing the large geometry devices.

In an example shown in Figure 23, a characteristic in which the slope of  $V_L$  becomes  $m''$  smaller than  $m'$  when the external supply voltage  $V_{CC}$  has reached  $V_0''$  is further added to the characteristic illustrated in Figure 17.

Figure 24 shows an embodiment of a practicable circuit therefor. This corresponds to an example in which the examples of Figures 7 and 9 coexist. The feature of the present circuit is that the embodiments of Figures 18 and 21 are combined thereby to increase and decrease the slope of  $V_L$  at the two points  $V_0'$  and  $V_0''$  respectively.

With the present circuit, the second and third control starting voltages  $V_0'$  and  $V_0''$  are respectively expressed by:

$$V_0' = V_0 + \left\{ \sum_{i=1}^{n'} V'_{th(i)} + V'_{th(l)} \right\} / (1 - m)$$

$$V_0'' = V_0' + \left\{ \sum_{i=1}^{n''} V''_{th(i)} + V''_{th(l)} + \sum_{i=1}^{n'} V'_{th(i)} + V'_{th(l)} - V_0' \right\} / m'$$

In addition, the slope  $m'$  is expressed by the ratio between the sum of the conductances of  $Q_0$  and  $Q_L'$  and the conductance of  $Q_L$ , while  $m''$  <sup>is expressed</sup> by the ratio between the sum of the conductances of  $Q_0$  and  $Q_L'$  and the sum of the conductances of  $Q_L$  and  $Q_L''$ .

Accordingly,  $V_0'$  and  $m'$  can be varied at will by  $n, n', \beta(0), \beta(l), \beta'(l), V_{th(i)}, V_{th(l)}, V'_{th(i)}$  and  $V'_{th(l)}$ , while  $V_0''$  and  $m''$  <sup>can be varied</sup> by  $n, n', n'', \beta(0), \beta(l), \beta'(l), \beta''(l), V_{th(i)}, V_{th(l)}, V'_{th(i)}, V'_{th(l)}, V''_{th(i)}$  and  $V''_{th(l)}$ .

This circuit protects small geometry devices from permanent breakdown in such a way that, even when  $V_{CC}$  has become higher than the withstand voltage limit  $V_0''$  of the devices due to <sup>some</sup> ~~any~~ fault of the external power source, it does not exceed a breakdown voltage  $V_B$ . That is, the slope  $m''$  of  $V_L$  for  $V_{CC}$  not smaller than  $V_0''$  is made gentler than the slope  $m'$  in the aging, whereby even when the external supply voltage  $V_{CC}$  has become  $V_0''$  or above, the output voltage  $V_L$  is prevented from exceeding the breakdown voltage (usually, higher than the withstand voltage limit) of the devices. This makes it possible to prevent the device breakdown even when the supply voltage has been raised abnormally by way of example.

Figure 25 shows an example in which the slope  $m'$  is made negative when the external supply voltage  $V_{CC}$  has exceeded  $V_0'$ .



Figure 26 shows an embodiment of a practical circuit therefor. These correspond to a concrete form of the example of Figures 11 and 12. The feature of the present circuit is that the drain of  $Q_1'$  in DCV2 is connected to the terminal 1 of the circuit shown in Figure 16, the drain of  $Q_\ell'$  to the terminal 2, and the source of  $Q_\ell'$  to the ground, whereby the conductance of  $Q_\ell'$  is controlled by  $V_{CC}$ , and besides, it is made greater than the conductance of  $Q_0$  so as to establish  $m' < 0$ .

With the present circuit, the second control starting voltage  $V_0'$  and the slope  $m'$  are expressed by the following on the assumption of  $\beta'(\ell) \gg \beta(0)$ :

$$V_0' = \sum_{i=1}^{n'} V'_{th(i)} + V'_{th(\ell)}$$

$$m' = 1 - \sqrt{\beta'(\ell)/\beta(0)}$$

Accordingly,  $V_0'$  and  $m'$  can be varied at will by  $n'$ ,  $V'_{th(i)}$ ,  $V'_{th(\ell)}$  and  $\beta'(\ell)/\beta(0)$ .

Figures 27 and 28 show a practicable example of the present circuit and examples of the characteristics thereof. All the threshold voltages of transistors are 1 (one) V, and  $V_G = V_{CC} + V_{th(0)}$  is held. In addition, numerals in parentheses indicate values obtained by dividing the channel widths by the channel lengths of the transistors. Figure 28 illustrates  $V_L$  with a parameter being the corresponding value  $W_\ell/L_\ell$  of  $Q_\ell'$ . By way of

example, the voltage in the ordinary operation is set at 5 V, and the aging voltage at 8 V.

This circuit consists in that the slope of the voltage at and above  $V_0$  in the characteristic shown in Figure 23 is made negative, thereby to intensify the aspect of the device protection of the circuit in Figure 24.

With this circuit, the breakdown due to the external application of a high voltage is perfectly prevented, and the power consumption in the integrated circuit does not exceed an allowable value. Thus, even when the instantaneous high voltage has been externally applied, the prevention of the breakdown of the devices is ensured.

Thus far, the voltage converters and their characteristics have been described. Next, the method of feeding the voltage converter with power will be described.

b In the above, the gate voltage of  $Q_0$  has been <sup>presumed to be</sup> ~~supposed~~  $V_{CC} + V_{th}$ . This has intended to simplify the computation and to clearly elucidate the characteristics of the circuits. Essentially, however, this voltage need not be <sup>limited</sup> ~~stuck~~ to  $V_{CC} + V_{th}$ , but may be chosen at will for the convenience of design.

b Figure 29(A) shows a practicable circuit which boosts the gate voltage  $V_G$  to above the supply voltage  $V_{CC}$  within the chip as stated with reference to Figure 15.

When a pulse  $\phi_i$  of amplitude  $V_{CC}$  from an oscillator OSC included within the chip rises from 0 (zero) V to  $V_{CC}$ , a node 4' having been previously charged to  $V_{CC} - V_{th}$  by  $Q_1'$  is boosted to  $2 V_{CC} - V_{th}$ .

In consequence, a node 4 becomes a voltage  $2 (V_{CC} - V_{th})$  lowered by  $V_{th}$  by means of  $Q_2'$ . Subsequently, when  $\phi_i$  becomes 0 V and a node 2 rises to  $V_{CC}$ , the node 4 is further boosted into  $3 V_{CC} - 2 V_{th}$ . Accordingly, a node 5 becomes a voltage  $3 (V_{CC} - V_{th})$  lowered by  $V_{th}$  by means of  $Q_2$ . Each of  $Q_2'$  and  $Q_2$  is a kind of diode, so that when such cycles are continued a large number of times,  $V_G$  becomes a D.C. voltage of  $3 (V_{CC} - V_{th})$ .  $V_G$  of higher voltage is produced by connecting the circuits CP1, CP2 in a larger number of stages. The reason why the two stages are comprised here, is as follows. Assuming  $V_{CC}$  to lower to 2.5 V and  $V_{th}$  to be 1 (one) V, one stage affords  $V_G = 2 (V_{CC} - V_{th})$ , and hence,  $V_G = 3$  V holds. Under this condition, however, the source voltage  $V_L$  of  $Q_0$  in Figure 15 becomes 2 V lower than  $V_{CC}$ . In contrast, when the two stages are disposed,  $V_G = 4.5$  V holds because of  $V_G = 3 (V_{CC} - V_{th})$ . Accordingly,  $V_L$  can be equalized to  $V_{CC}$ , so that  $V_L = V_{CC}$  can be established below  $V_O$  as in Figure 15. Conversely, however, as  $V_{CC}$  becomes a higher voltage,  $V_G$  <sup>may</sup> ~~is more feared to~~ become an excess voltage <sup>which can</sup> ~~and to~~ break down the associated transistors.

b Therefore, <sup>some</sup> ~~any~~ circuit for limiting  $V_G$  is required on the high voltage side of  $V_{CC}$ .

Figure 30 shows an example in which  $V_G \simeq 3 (V_{CC} - V_{th})$  is held as a high voltage on the low voltage side of  $V_{CC}$ , and besides,  $V_{CC} + 2 V_{th}$  is held on the high voltage side of  $V_{CC}$  in order to protect the associated transistors. Here, any of the circuits thus far described, for example, the whole circuit in Figure 16, 18, 20, 22, 24 or 26, is indicated by LML as the load of  $V_G$ . A protection circuit CL1 is such that, when  $V_G$  is going to exceed  $V_{CC} + 2 V_{th}$ , current flows through  $Q_1$  and  $Q_2$ , so  $V_G$  results in being fixed to  $V_{CC} + 2 V_{th}$ . With the present circuit,  $V_{CC}$  at which CL1 operates ranges from  $3 (V_{CC} - V_{th}) = V_{CC} + 2 V_{th}$  to  $V_{CC} = 5/2 V_{th}$ .

Figure 31 shows a practicable circuit of the inverter 1 or 2 in Figure 29(A). An output pulse  $\phi_0$  is impressed on the circuit CP1 or CP2.

While the oscillator OSC can be constructed as a circuit built in the chip, Figure 32 shows an example utilizing a back bias generator which is built in the chip in order to apply a back bias voltage  $V_{BB}$  to a silicon substrate. The advantage of this example is that the oscillator need not be designed anew, which is effective for reducing the area of the chip. In general, when  $V_L$  is applied to respective transistors

with  $V_{BB}$  being 0 (zero) V, the threshold voltages  $V_{th}$  of the respective transistors are not normal values. Therefore, an excess current flows, or stress conditions on the transistors become severe, so the transistors can break down. In contrast, when this circuit is used,  $V_{BB}$  is generated upon closure of a power source, and  $V_L$  is generated substantially simultaneously, so that the operations of respective transistors are normally executed.

Next, practicable embodiments of buffer circuits will be described. As the load of the voltage converter, there is sometimes disposed a load of large capacity or of great load fluctuation. In this case, such heavy load needs to be driven through a buffer circuit of high driving ability. As a realizing expedient therefor, the ordinary method is considered in which the load is driven through a single transistor of high driving ability, namely, great W/L as shown in Figure 33. With this method, however, the performance degrades because a voltage drop of  $V_{th}$  arises on the low voltage side of  $V_{CC}$  as shown in Figure 34. Figure 35 shows a practicable example of the buffer circuit which has a high driving ability without the  $V_{th}$  drop. When a voltage  $V_{pp}$  is made greater than  $V_L + V_{th}$  and a resistance  $R_p$  is made much higher than the equivalent "on" resistance of a

transistor  $Q_1$ , the gate voltage of a transistor  $Q_2$  becomes  $V_L + V_{th}$ . Accordingly, the source voltage  $V_{L1}$  of  $Q_2$  equalizes to  $V_L$ . When the W/L of  $Q_2$  is made great, the desired buffer circuit is provided. Here,  $V_L$  becomes  $V_{CC}$  on the low voltage side of  $V_{CC}$ , so that  $V_{PP}$  must be at least  $V_{CC} + V_{th}$ . As a circuit therefor, the circuit shown in Figure 29(A) is usable. Regarding connection, the node 5 of the circuit in Figure 29(A) may be connected to the drain of  $Q_1$  in a regulator in Figure 35. Here, in order that the effective output impedance as viewed from the node 5 may be made sufficiently higher than the equivalent "on" resistance of  $Q_1$  of the circuit in Figure 35, the value of the W/L of  $Q_2$  or the value of  $C_B$  in Figure 29(A) or the oscillation frequency of OSC may be properly adjusted by way of example.

As to some loads, it is necessary to apply  $V_L$  to the drain of a transistor constituting a part of the load and to apply  $V_L + V_{th}$  to the gate thereof, so as to prevent the  $V_{th}$  drop and to achieve a high speed operation. Figure 36 shows an embodiment therefor. The circuit  $LM_1$  is, for example, the circuit in Figure 16, and the voltage  $V_{L1}$  equalizes to  $V_L$  as stated before. In addition, the gate voltage of  $Q_4$  is  $V_L + 2 V_{th}$ . Therefore,  $V_{L2}$  becomes  $V_L + V_{th}$ . Here, transistors  $Q_6$  and  $Q_7$  serve to prevent unnecessary charges from remaining in  $V_{L1}$ .

at the transient fluctuation of  $V_{CC}$ .  $Q_6$  and  $Q_7$  are connected into LML as shown in the figure so as to operate at  $V_{CC}$  of at least  $V_O$  and at  $V_{CC}$  of at least  $V_O - V_{th}$ . Here, the ratio  $W/L$  of  $Q_6$ ,  $Q_7$  is selected to be sufficiently smaller than that of  $Q_2$ , to minimize the influence of the addition of  $Q_6$ ,  $Q_7$  on  $V_L$ . It has been previously stated that  $Q_7$  operates in the region not greater than  $V_O$ . Since  $Q_2$  and  $Q_4$  are in the operating states of unsaturated regions ( $V_{GS} - V_{th} \geq V_{DS}$ ,  $V_{GS}$ : gate-source voltage,  $V_{DS}$ : drain-source voltage) in the region not greater than  $V_O$ , surplus charges are discharged to  $V_{CC}$  through  $Q_2$ ,  $Q_4$ , and hence,  $Q_7$  is unnecessary in principle. However, when  $V_{CC}$  is near  $V_O$ , the "on" resistances of  $Q_2$ ,  $Q_4$  increase unnecessarily, and it is sometimes impossible to expect the effects of these transistors. Accordingly,  $Q_7$  is added, whereby stable values of  $V_{L1}$  can be obtained in a wide range from the region ( $V_O - V_{th}$ ) where  $V_{CC}$  is not greater than  $V_O$ , to the region where  $V_{CC}$  is greater than  $V_O$  and where the converter is normally operating.

The function of  $Q_5$  is that, when  $V_{L1}$  is going to fluctuate negatively relative to  $V_{L2}$ , current flows to  $Q_5$  so as to keep the difference of  $V_{L2}$  and  $V_{L1}$  constant. In addition, in the present embodiment, the example of  $V_L$  and  $V_L + V_{th}$  has been stated. However, when the pairs of  $Q_1$ ,  $Q_2$  or the pairs of  $Q_3$ ,  $Q_4$  are connected

in cascade, a voltage whose difference from  $V_{L1}$  becomes an <sup>multiple</sup> integral ~~times~~ of  $V_{th}$  can be generated.

A circuit shown in Figure 37 is another buffer circuit which is connected to the output stage of the circuit of Figure 35 or 36 in order to <sup>further</sup> ~~more~~ enhance the driving ability of the buffer circuit of Figure 35 or 36. By connecting such <sup>a</sup> buffer circuit of higher driving ability, a large load capacity can be driven. First,  $V_{L1}$  becomes  $V_{L1} + 2 V_{th}$  and  $V_{L1} + V_{th}$  at respective nodes 4 and 2. Eventually, however, it is brought into  $V_{DP}$  being the level of  $V_{L1}$  at a node 5 by  $Q_4$ . ~~Problematic here is the load-driving ability of  $Q_4$  which serves to charge a large capacitance  $C_D$  in the load LCL at high speed.~~ In order to enhance the ability, the node 2 being the gate of  $Q_4$  needs to be boosted in a time zone for charging the load. Transistors ~~therefor are~~  <sup>$Q_6 - Q_{11}$</sup> , and capacitors are  $C_1$  and  $C_2$ . A node 6 discharged by  $Q_{13}$  owing to <sup>the state</sup> "on" of  $\phi_2$  is charged by  $Q_{12}$  and  $Q_4$  when the next  $\phi_1$  <sup>is in the</sup> ~~is on~~ "on" state. At this time, the node 2 being at  $V_{L1} + V_{th}$  and a node 3 being at  $V_{L1}$  are boosted by the <sup>state</sup> "on" of  $\phi_1$ . Consequently, the conductances of  $Q_{10}$ ,  $Q_{11}$  increase, so that the boosted voltage of the node 2 is discharged to the level of  $V_{L1} + V_{th}$  by  $Q_{10}$ ,  $Q_{11}$ . Here, when the boosting time is made longer than the charging time of  $C_D$  based on  $Q_4$ ,  $Q_{12}$ , the capacitor  $C_D$  is charged



<sup>rapidly</sup>  
~~fast~~. The transistor  $Q_6$  cuts off the nodes 3 and 1  
 when the node 3 is boosted by  $\phi_1$  <sup>control signal</sup>. When  $\phi_2$  <sup>control signal</sup> is "on",  
 $Q_7 - Q_9$  turn "off" subject to the condition of  $V_{L1} \leq 3 V_{th}$ ,  
 so that  $Q_{11}$  has its gate rendered below  $V_{th}$  to turn  
 "off". Accordingly, no current flows through  $Q_3$ ,  $Q_{10}$   
 and  $Q_{11}$ , so that the power consumption can be rendered  
 low. In addition, in order to reduce the power consumption  
 in the case of  $V_{L1} > 3 V_{th}$ , the "on" resistance of  $Q_6$   
 may be increased to lower current. The voltage of the  
 node 3 at this time becomes a stable value of approximately  
 $3 V_{th}$ . Thus, the boosting characteristic of the node  
 3 is also stabilized, with the result that the operation  
 of the whole circuit can be stabilized.

Here, since the sources and gates of  $Q_7$  and  $Q_{10}$   
 are connected in common, the conditions of biasing the  
 gates are quite equal. Accordingly, when

$$\frac{\text{capacitance of node } \overset{3}{2}}{(W/L) Q_7} = \frac{\text{capacitance of node } \overset{2}{3}}{(W/L) Q_{10}}$$

is held in advance, the boosting characteristics of  
 the nodes 2, 3 can be made quite equal, so the circuit  
 design can be facilitated advantageously. That is, one  
 merit of the present embodiment consists in that the  
 boosting characteristic of the node 2 can be automatically  
 controlled with the boosting characteristic of the node  
 3. In this way, the D.C. path from the node 2 to  $V_{SS}$

in the case of performing no boosting can be relieved, and it becomes possible to lower the power consumption.

Here,  $Q_5$  has the function of discharging the surplus charges of the node 2 when  $Q_{10}$  is "off".

As regards the embodiment of Figure 37, various modifications can be considered. While the drain of  $Q_6$  in Figure 37 is connected to  $V_{L1}$  in order to stabilize the boosting characteristics of the nodes 2, 3 to the utmost, it can also be connected to  $V_{CC}$  so as to relieve a burden on  $V_{L1}$ . Likewise, while  $Q_{10}$  subject to the same operating condition as that of  $Q_7$  is disposed in order to stabilize the boosting characteristics of the nodes 2, 3, it may well be removed into an arrangement in which the nodes 2 and 9 are directly connected, with the source of  $Q_7$  and the node 9 disconnected. Since, in this case, the relationship of  $Q_9$  and  $Q_{11}$  is in the aforementioned relationship of  $Q_7$  and  $Q_{10}$ , the boosting characteristics can be similarly designed, and the occupying area of the circuit can be effectively reduced. Further, the 3-stage connection arrangement of  $Q_7$ ,  $Q_8$  and  $Q_9$  is employed here. This is a consideration for efficiently forming the circuit in a small area by utilizing a capacitance  $C_2$  (for example, the capacitance between the gate of a MOST and an inversion layer formed between the source and drain thereof, known from ISSCC 72 Dig. of Tech. Papers, p. 14, etc.) for the

reduction of the power consumption described above. That is, in order to use the inversion layer capacitance, the gate voltage to be applied needs to be higher by at least  $V_{th}$  than the source and drain. Accordingly, in case of forming  $C_2$  by the use of a MOST of low  $V_{th}$  or an ordinary capacitor, it is also possible to reduce the connection number of  $Q_7 - Q_9$  to two or one.

The buffer circuit as shown in Figure 37 is indispensable especially to the LSI systems as shown in Figures 1 and 2. In general, the voltage converter for generating  $V_L$  in Figure 1 or 2 is desired to have an especially high ability of supplying current because the circuit current in the circuit A, A' or B flows toward the ground. Accordingly, when the whole circuit including the circuit of Figure 37 thus far described is regarded as the voltage converter of Figure 1 or 2, it is applicable to general LSIs.

With the embodiments stated above, when the actual circuit of Figure 18 which is diode-connected as shown in Figure 27 is operated at  $V_{CC}$  of or above  $V_O$  as shown in Figure 17, current flows through  $Q_1' - Q_5'$  (Figure 27) to increase the power consumption. This increase of the power consumption poses a problem in case of intending to back up the LSI power source, namely, the externally applied supply voltage with a battery. More specifically,

in an apparatus wherein the ordinary external power source is backed up by a battery when turned "off"; when the power consumption of the LSI itself is high, the period of time for which the power source is backed up is limited because the current capacity of the battery is small. Therefore, with a measure wherein  $V_{CC}$  to be applied by the battery is set at below  $V_0$  during the time interval during which the battery is operated for backup, no current flows through  $Q_1' - Q_S'$ , and hence, the period of time for which the power source can be backed up can be extended to that extent. Alternatively, the number of stages of  $Q_1' - Q_S'$  (Figure 27) can be determined so as to establish  $V_0$  which is greater than  $V_{CC}$  being the battery supply voltage in the case of the backup.

The supply voltage  $V_{CC}$  in the ordinary operation can be selected at  $V_{CC} < V_0$  besides at  $V_{CC} > V_0$ . Since this permits no current to flow through  $Q_1' - Q_S'$  under the ordinary  $V_{CC}$  condition, the power consumption can be lowered. Another merit is that design is facilitated because the circuit can be designed <sup>while</sup> ~~whilst~~ avoiding a region where the relation of  $V_{CC}$  and  $V_L$  becomes a polygonal line. More specifically, when the polygonal region is used, <sup>an</sup> ~~the~~ imbalance of characteristics concerning  $V_{CC}$  arises between a circuit directly employing  $V_{CC}$  and

b  
a part of a certain circuit employing  $V_L$  by way of example, so that the operation sometimes becomes unstable. When  $V_{CC} < V_0$  <sup>holds</sup> ~~is held~~, this drawback can be eliminated.

In the above, the practicable embodiments have been described in which the voltage converters are constructed of MOS transistors. These are examples which chiefly employ MOS transistors of positive threshold voltages  $V_{th}$ , namely, of the enhancement mode. Needless to say, however, it is also possible to employ a MOS transistor of negative  $V_{th}$ , namely, of the depletion mode as disclosed in Figure 16 of Japanese Patent Application No. 56-168698. For example, in the embodiment of Figure 16, in order to establish  $V_L = V_{CC}$  in the region of  $V_{CC} \leq V_0$  as illustrated in the characteristic of Figure 15, the gate voltage of  $Q_0$  needs to be  $V_G \geq V_{CC} + V_{th(0)}$ , and it has been stated that the circuit of Figure 29(A) may be used as the  $V_G$  generator therefor. In this regard, the circuit can be <sup>further</sup> ~~more~~ simplified by employing the MOS transistor <sup>such</sup> of the depletion mode. Figure 39 shows a practicable embodiment. It differs from the circuit of Figure 16 in that  $Q_0$  is replaced with the depletion mode MOS transistor  $Q_0'$ , the gate of which is connected to the terminal 2. With this measure, since the  $V'_{th(0)}$  of  $Q_0'$  is negative,  $Q_0'$  is in the "on" state at all times, and the desired characteristic illustrated in Figure 15 can be realized

b

without employing the  $V_G$  generator as shown in Figure 29(A). With the present embodiment, not only the circuit arrangement can be simplified as stated above, but also the merit of attaining a stable characteristic is achieved because current  $I(Q_0')$  to flow through  $Q_0'$  becomes a constant current determined by  $\beta'(0)$  (channel conductance) and  $V'_{th(0)}$  (threshold voltage) as  $I(Q_0') = \frac{\beta'(0)}{2} \cdot V'_{th(0)}{}^2$ . Although the present embodiment has exemplified Figure 16, it is applicable as it is by substituting  $Q_0'$  for  $Q_0$  in any other embodiment and connecting its gate to the terminal 2 as in the present embodiment.

Figure 40 shows an embodiment in which a buffer circuit is constructed using a single depletion-mode MOS transistor, while Figure 41 shows the characteristic thereof. Although the present embodiment is the same in the circuit arrangement as the foregoing embodiment of Figure 33, it differs in that the MOS transistor is changed from the enhancement mode into the depletion mode. As shown in Figure 41, the output  $V_L'$  of the present buffer circuit bends from a point P at which the difference of  $V_{CC}$  and  $V_L$  equalizes to the absolute value  $|V_{thD}|$  of the threshold voltage  $V_{thD}$  of the MOS transistor, and it thereafter becomes a voltage which is higher than  $V_L$  by  $|V_{thD}|$ . Accordingly,  $V_L$  may be set lower than a desired value by  $|V_{thD}|$ . The present

embodiment has the simple circuit arrangement, and can meritoriously eliminate the problem, as in the characteristic of the embodiment of Figure 33 illustrated in Figure 34, that only the output lower than  $V_{CC}$  by  $V_{th}$  can be produced in the range of  $V_{CC} \leq V_O$ .

As set forth above, the present invention can provide, in an integrated circuit having small geometry devices, an integrated circuit which has a wide operating margin even against the fluctuations of an external supply voltage in an ordinary operation and which can apply a sufficient aging voltage.